BROADBAND COAXIAL CHOKE COUPLING DESIGN

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SUMMARY

Equations and curves are presented to predict the frequency bandwidth of coaxial choke couplings in terms of the choke parameters. Choke couplings discussed are those applicable to rotary joints and dc isolation units.

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I. INTRODUCTION

Coaxial choke type rotary joint designs have been discussed by Ragan,[1] and many have been built following his presentation. Recently Muehe[2] discussed a method to widen the bandwidth of coaxial choke-type rotary joints by reducing the characteristic admittance of the transmission line for a quarter wavelength on each side of the chokes. Muehe's discussion was based on the analogous case of broadbanding short-circuited quarter-wavelength stubs in parallel with the transmission line, by changing the characteristic impedance of the line on each side of the stub for a distance of one-quarter wavelength.

The broadbanding of coaxial choke couplings under the present discussion is not based on a change in transmission line impedance, but is based on an extension of the conventional methods.

As outlined by Ragan, broadbanding of choke couplings may be done by displacing the outer and inner conductor chokes along the transmission line by one-quarter wavelength. The purpose of this paper is to present general equations and curves relating to the VSWR, the characteristic impedance of the choke sections, and the spacing of the two chokes. From these curves, one can predict the bandwidth of a rotary joint design.

In addition to the design of rotary joints, wideband dc isolation units can be built using the information presented herewith. Dc isolation units are necessary whenever blocking of dc on both the inner and outer conductors is desired. Wideband dc isolation units have been built[3] using the design described.

II. EQUATIONS FOR CHOKE COUPLING

A conventional coaxial rotary joint is shown in Figure 1. To prevent radiation losses due to the outer conductor choke and to provide a means of placing a bearing at a low current point, the external choke section,

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1. G. L. Ragan, "Microwave Transmission Circuits", M. I. T. Rad. Lab. Ser., McGraw-Hill Book Co., Inc., New York, New York, vol.9, p. 407; 1948.

2. C. E. Muehe, "Quarter-Wave Compensation of Resonant Discontinuities", Trans. I. R. E., vol. MTT-7, pp. 296-297; April 1959.

3. H. E. King, unpublished data and notes, built at Ramo-Wooldridge, a division of Thompson Ramo Wooldridge Inc.

of characteristic impedance Z_{o3}, is added. In most practical cases the characteristic impedance Z_{o3} is made as high as possible and usually much greater than the characteristic impedance Z_{o2} of the outer conductor choke.

For simplification in this analysis, Z_{o3} was assumed to be infinite. Also, both the choke sections were $\lambda/4$ long at the center frequency, and its characteristic impedances Z_{o1} and Z_{o2} were assumed to be equal; thus, the choke input impedances were equal, or $Z_1 = Z_2$. The characteristic impedance, Z_o, of the transmission line was normalized to 1.

The ABCD matrix of the two chokes displaced by the length ℓ along a lossless transmission line is

$$\begin{bmatrix} 1 & -jZ_{o1}\cot\beta\ell_1 \\ 0 & 1 \end{bmatrix} \begin{bmatrix} \cos\beta\ell & j\sin\beta\ell \\ j\sin\beta\ell & \cos\beta\ell \end{bmatrix} \begin{bmatrix} 1 & -jZ_{o1}\cot\beta\ell_1 \\ 0 & 1 \end{bmatrix}, \tag{1}$$

where $\beta\ell_1$ is the electrical length of the choke sections and $\beta\ell$ is the electrical spacing between the two chokes. The final matrix when multiplied through is

$$\begin{bmatrix} \cos\beta\ell + Z_{o1}\cot\beta\ell_1\sin\beta\ell & -2jZ_{o1}\cot\beta\ell_1\cos\beta\ell - jZ_{o1}^2\cot^2\beta\ell_1\sin\beta\ell + j\sin\beta\ell \\ j\sin\beta\ell & \cos\beta\ell + Z_{o1}\cot\beta\ell_1\sin\beta\ell \end{bmatrix} \tag{2}$$

The insertion loss is given by[4]

$$\begin{aligned} L &= 10\log_{10}\left\{1 + 1/4\left[(A - D)^2 - (B - C)^2\right]\right\} \\ &= 10\log_{10}\left\{1 - 1/4\left[-j(2Z_{o1}\cot\beta\ell_1\cos\beta\ell + Z_{o1}^2\cot^2\beta\ell_1\sin\beta\ell)\right]^2\right\} \\ &= 10\log_{10}\left\{1 + \frac{|K|^2}{4}\right\} \quad . \end{aligned} \tag{3}$$

Next, let $\beta\ell_1 = \frac{\pi}{2} + \phi$ and $\beta\ell = n\beta\ell_1 = n\left(\frac{\pi}{2} + \phi\right)$.

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4. R. M. Fano and A. W. Lawson, "Microwave Transmission Circuits", M.I.T. Rad. Lab. Ser., McGraw-Hill Book Co., Inc., New York, New York, vol. 9, p. 551; 1948.

Then $|K|$ is written as

$$|K| = 2Z_{o1} \tan\phi \cos n\left(\frac{\pi}{2} + \phi\right) - Z_{o1}^{2} \tan^{2}\phi \sin n\left(\frac{\pi}{2} + \phi\right), \tag{4}$$

where ϕ = the difference in electrical length from the center frequency choke length of $\frac{\pi}{2}$. $|K|$ is related to the coaxial transmission line voltage-standing-wave ratio, S, by

$$|K| = \frac{S - 1}{\sqrt{S}} \quad . \tag{5}$$

Bandwidth is arbitrarily defined by the condition when the transmission line VSWR is 1.1 (insertion loss less than 0.01 db) or in symbolic form, the bandwidth is

$$\frac{f_2}{f_1} = \frac{90 + |\phi_2|}{90 - |\phi_1|} \quad . \tag{6}$$

III. DISCUSSION OF CURVES

A graph of the input VSWR to the transmission line is shown in Figure 2 for the case when Z_{o1} = .0324 for various conditions of n. When n = 0, the input terminals of the chokes are located on the same transverse plane. The VSWR is calculated from

$$|K|_{n=0} = 2Z_{o1} \tan\phi. \tag{7}$$

When the chokes are displaced by $\lambda/4$ at the center frequency, then n = 1, or equation (4) is reduced to

$$|K|_{n=1} = \left(2Z_{o1} + Z_{o1}^{2}\right) \tan\phi \sin\phi. \tag{8}$$

When n = 1, a zero derivative exists at the origin, a condition considered to be the maximally flat case.

The curve for n = 2, or when the input choke terminals are separated by $\lambda/2$ at the center frequency, shows a wider bandwidth. $|K|$ is written as

$$|K|_{n=2} = 2Z_{o1} \tan\phi \cos^{2}\phi - 2\left(Z_{o1} + Z_{o1}^{2}\right) \tan\phi \sin^{2}\phi. \tag{9}$$

When the chokes are separated by $\lambda/4$ at the center frequency, or n = 1, the frequency bandwidth ratio vs. Z_{o1} is plotted in Figure 3, where the band edge limits were determined by a voltage-standing-wave ratio of 1.1.

Figure 4(a) is a curve of frequency bandwidth ratio vs Z_{o1} for n = 2. The peak voltage-standing-wave-ratio within the band limits is plotted in Figure 4(b).

Where Z_{o1} = .0324 and n = 2, the bandwidth for a VSWR of less than 1.1:1 is 6.2:1. With the same choke impedance except that n = 2.67 ($\beta\ell = 240^\circ$ at the center frequency), a still wider frequency bandwidth of 8.15:1 is theoretically feasible as illustrated in Figure 5. Note that the curve is unsymmetrical and the peak within the band is slightly higher. In most practical cases, there should not be any detrimental effects. With any given value of Z_{o1} an optimum spacing between chokes can be determined to give the widest frequency bandwith.

IV. OTHER CONSIDERATIONS

For extremely wide bandwidth choke couplings, the finite value of Z_{o3} should be considered. The solid line curve of Figure 5 represents the condition when the external choke impedance $Z_{o3} = \infty$, while the broken line curves represent two finite values of Z_{o3}.

The dashed line curve of Figure 5 shows the reduction in bandwidth if the normalized charateristic impedance of Z_{o3} is 0.6 ohms. For a normalized characteristic impedance Z_{o3} of 1.5 ohms, the dash-dot curve indicates an improvement in the VSWR response. Note that there is no major increase in VSWR due to the finite value of Z_{o3} provided $|\phi| < 70^\circ$. Usually only for extremely wide bandwidth choke coupling designs (when $|\phi| > 70^\circ$) is it necessary to calculate the effects of Z_{o3} on the VSWR.

If the external diameter is a limiting factor preventing a large value of Z_{o3} to be selected, an effective increase in the external choke impedance can be obtained by making it $\lambda/2$ long at the mid-frequency as illustrated in Figure 6. Now the impedance $Z_3 + Z_4$ terminates the outer conductor choke Z_{o2}, instead of only Z_3.

In a practical design, the slight reduction in bandwidth due to the finite impedance of the external choke can be compensated by lowering the characteristic impedance Z_{o2}. In a practical case, Z_{o2} can be made smaller than Z_{01}.

Basically, the wide bandwidth of choke couplings is obtained by the use of low values of the choke's characteristic impedances, Z_{o1} and Z_{o2}. By the use of teflon insulation, choke characteristic impedances of less than 1.5 ohm is readily obtained. The increase in dissipation losses because of the use of teflon dielectric is compensated for by the shorter physical length* of choke that is required.

Figure 7 is a photograph of a dc isolation unit built[3] for a frequency range of 50 mc to 900 mc (insertion loss less than 0.5 db) using the design described herewith.

The above curves and analysis should be helpful for the design of wide band coaxial rotary joints or dc isolation units. Prediction of the final performance with various choices of parameters is possible.

*See Appendix

APPENDIX

To demonstrate the effects of the choke losses due to the use of teflon dielectric, the following relations are presented. The input resistance to a resonant transmission line with a relative dielectric constant of 1, is expressed as

$$R_{in_{\epsilon = 1}} = Z_o \alpha \ell = \frac{R\ell}{2} + \frac{GZ_o^2 \ell}{2} , \qquad \text{(A1)}$$

where $\alpha = \frac{R}{2Z_o} + \frac{GZ_o}{2}$.

Assuming that the dielectric constant is increased to ϵ_r, the characteristic impedance Z_o and length ℓ, are both reduced by $1/\sqrt{\epsilon_r}$. A dielectric filled resonant line, assuming the same dielectric dissipation factor of equation (A1) will have the input resistance

$$R_{in_{\epsilon = \epsilon_r}} = \frac{R\ell}{2\sqrt{\epsilon_r}} + \frac{GZ_o^2 \ell}{2 \epsilon_r^{\frac{3}{2}}} . \qquad \text{(A2)}$$

Note that the copper loss term is reduced by $1/\sqrt{\epsilon_r}$. Furthermore, the dielectric loss term is reduced by a factor of $\frac{1}{\epsilon_r^{\frac{3}{2}}}$.

In actuality, the dielectric loss is increased from zero to a finite value when converting an air dielectric filled choke to a teflon dielectric filled choke. However, the loss tangent of teflon is very small, Z_o is small, and the dielectric loss term is also reduced by $\frac{1}{\epsilon_r^{\frac{3}{2}}}$. Thus, in a practical case, instead of increased dissiptation losses due to the use of teflon dielectric, the losses are reduced.